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(54) Title: PRACTICAL SPACE-TIME RADIO METHOD FOR CDMA COMMUNICATION CAPACITY ENHANCEMENT

(57) Abstract

A practical way to enhance signal quality (carrier to interference, C/I) in both up and downlink of wireless point to multi–point CDMA service implements basic radio direction finding techniques to allow for optimal diversity combining in an antenna array employing large number of elements. This approach is facilitated through the use of very small bit counts arithmetic and capitalizing on finite alphabet signal structure (Walsh symbols, for example in IS–95 CDMA) or a known training sequence. Alternate implementations can use floating point data representations. The method facilitates ASIC implementation, thereby enabling distributed processing to achieve the required computation practicality. The method utilizes the uplink channel data to determine the downlink spatial structure (array beams) to enhance downlink C/I and hence increase downlink capacity. The preferred embodiment is optimized to IS–95, however any signal that has either a finite alphabet or a training sequence built in can utilize the same idea. The use of the known signal structure facilitates simple array response vector determination and eliminates the necessity f r covariance matrix calculation and analysis. Hence, this approach can be utilized for GSM and TDMA wireless air–interfaces as well.

PRACTICAL SPACE-TIME RADIO METHOD FOR
CDMA COMMUNICATION CAPACITY ENHANCEMENT

5

BACKGROUND

Field of the invention

10   The present invention relates to wireless
communication systems. More specifically, the
invention relates to methods for enhancement of

15   wireless communication performance by exploiting the
spatial domain, and practical systems for implementing
such methods.

Related Art

20   Due to the increasing demand for wireless
communication, it has become necessary to develop
techniques for more efficiently using the allocated
frequency bands, i.e., increasing the capacity to
communicate information within a limited available

25   bandwidth. In conventional low capacity wireless
communication systems, information is transmitted from
a base station to subscribers by broadcasting
omnidirectional signals on one of several predetermined
frequency channels. Similarly, the subscribers

30   transmit information back to the base station by
broadcasting similar signals on one of the frequency
channels. In this system, multiple users independently
access the system through the division of the frequency
band into distinct sub-band frequency channels. This

35   technique is known as frequency division multiple
access (FDMA).

A standard technique used by commercial wireless
phone systems for increasing capacity is to divide the
service region into spatial cells. Instead of using
just one base station to serve all users in the region,
5　　a collection of base stations is used to independently
service separate spatial cells. In such a cellular
system, multiple users can reuse the same frequency
channel without interfering with each other, provided
the users access the system from different spatial
10　　cells. The cellular concept, therefore, is a simple
type of spatial division multiple access (SDMA).

In the case of digital communication, additional
techniques can be used to increase capacity. A few
well-known examples are time division multiple access
15　　(TDMA) and code division multiple access (CDMA). TDMA
allows several users to share a single frequency
channel by assigning their data to distinct time slots.
CDMA is normally a spread-spectrum technique that does
not limit individual signals to narrow frequency
20　　channels but spreads the signals throughout the
frequency spectrum of the entire band. Signals sharing
the band are distinguished by assigning different
orthogonal digital code sequences or spreading signals
to each signal. CDMA has been considered the most
25　　promising method among the various air-interfaces in
the industry, as shown by theoretical analysis (See,
for example, Andrew J. Viterbi, *CDMA Principles of
Spread Spectrum Communications*, and Vijay K. Garg et
al., *Applications of CDMA in Wireless/Personal*
30　　*Communications*.

Despite the promise of CDMA, practical issues such
as power control speed and inter-base station
interference considerably limited system effectiveness
in the initial phase of CDMA implementation. CDMA-
35　　based system capacity depends very much on the ability
to provide for very accurate power control, but in a

mobile environment, the signal may fluctuate too fast
for the system to control. Unfortunately, mobile
wireless environments are often characterized by
unstable signal propagation, severe signal attenuation
5   between the communicating entities, and co-channel
interference by other radio sources. Moreover, many
urban environments contain a significant number of
reflectors (such as buildings), causing a signal to
follow multiple paths from the transmitter to the
10 receiver. Because the separate parts of such a
multipath signal can arrive with different phases that
destructively interfere, multipath can result in
unpredictable signal fading. Furthermore, fast fading,
which is created by the combination of multipath
15 components of a signal being reflected from various
elements ("scatterers") in the neighborhood
("scattering zone") of a moving transmitter with random
phases, is considered to be a major issue in wireless
communication. The destructive combining at the
20 receiving antenna produces time varying signal levels
with a power density function characterized by a
Rayleigh distribution. Thus, the received power
experiences "deeps" or nulls at various times that can
cause significant errors in the transmitted information
25 (characterized by "burst bit errors" in digital
communication). In addition to fading, inter-base
station interference can cause significant system
performance degradation when radiated power is
increased in order to provide service to shadowed
30 areas.
     Modern communication systems reduce the fading
effects by interleaving the transmitted data and de-
interleaving the received data, with the addition of
proper error correction techniques. In addition,
35 utilizing spatial diversity is a very common method for
mitigating fading, e.g., a signal received at two

sufficiently spaced antennas (10 wavelengths or more)
has a small correlation in the received power vs. time
(power/time) function. Hence, most point-to-multipoint
communication systems utilize spatial diversity
5   combining to reduce fading effects. In most cases, the
receiver either selects the antenna with the stronger
signal power ("switching diversity") or combines two
antenna outputs after compensating for the phase and
amplitude difference ("maximum ratio combining").
10   Spread spectrum direct sequence systems (such as
IS-95) provide for additional fading mitigation by time
diversity, i.e., multipath can be separated by time due
to signal bandwidth and its associated auto-correlation
function. If multipath components arrive with
15   sufficient time spacing, their power/time functions are
not correlated. In IS-95, the RAKE receiver provides
for a plurality of demodulators ("fingers"), each
assigned to a different time of signal arrival.
Typically, the number of demodulating channels is four
20   at the base station. If the arriving signal multipath
has significant delay spread (e.g., several
microseconds), the system can successfully assign
different "fingers" to the incoming multipath
components and provide for excellent fading mitigation.
25   In most cases, however, the delay spread is not
sufficient to provide for time diversity (especially in
suburban areas), and the majority of fading mitigation
is still provided by spatial diversity and coding.
Since current base stations employ only two antennas
30   per sector, only two "fingers" are usually active.
Recently, considerable attention has focused on
ways to increase wireless system performance by further
exploiting the spatial domain. It is well recognized
that SDMA techniques, in principle, could significantly
35   improve the CDMA-based network performance. These
techniques have varying degrees of sophistication and

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complexity.   Currently proposed approaches are either
simple but not very effective or effective but too
complex for practical implementation.

One well-known SDMA technique is to provide the
5   base station with a set of independently controlled
directional antennas, thereby dividing the cell into
separate sectors, each controlled by a separate
antenna. As a result, the frequency reuse in the
system can be increased and/or co-channel interference
10   can be reduced. Instead of independently controlled
directional antennas, this technique can also be
implemented with a coherently controlled antenna array.
Using a signal processor to control the relative phases
of the signals applied to the antenna elements,
15   predetermined beams can be formed in the directions of
the separate sectors. Similar signal processing can be
used to selectively receive signals only from within
the distinct sectors. These simple sectoring
techniques, however, only provide a relatively small
20   increase in capacity.

U.S. Patent No. 5,563,610 discloses a method for
mitigating signal fading due to multipath in a CDMA
system. By introducing intentional delays into
received signals, non-correlated fading signal
25   components can be better differentiated by the RAKE
receiver. Although this diversity method can reduce
the effects of fading, it does not take advantage of
the spatial domain and does not directly increase
system capacity. Moreover, this approach, which
30   combines angular and time diversity using a fixed beam
configuration, is not effective since either the beam
outputs are significantly different in level or they
are similar in level but highly correlated. If two
signal parts are arriving from similar direction, they
35   are passing through one beam and thus are non-
differentiable. If the signal parts are arriving

between beams, on the other hand, the levels are
similar, but they are well correlated.

   More sophisticated SDMA techniques have been
proposed that could dramatically increase system
5 capacity. For example, U.S. Pat. No. 5,471,647 and
U.S. Pat. No. 5,634,199, both to Gerlach et al., and
U.S. Pat. No. 5,592,490 to Barratt et al. disclose
wireless communication systems that increase
performance by exploiting the spatial domain. In the
10 downlink, the base station determines the spatial
channel of each subscriber and uses this channel
information to adaptively control its antenna array to
form customized narrow beams. These beams transmit an
information signal over multiple paths so that the
15 signal arrives to the subscriber with maximum strength.
The beams can also be selected to direct nulls to other
subscribers so that co-channel interference is reduced.
In the uplink, the base station uses the channel
information to spatially filter the received signals so
20 that the uplink signal is received with maximum
sensitivity and distinguished from the signals
transmitted by other subscribers. Through selective
power delivery by intelligent directional beams, the
inter-base station interference and the carrier-to-
25 interference (C/I) ratio at the base station receivers
can be reduced.

   The biggest issue in adaptive beamforming is how
to quickly estimate the wireless air channel to allow
for effective beam allocation. In the uplink, there
30 are known signal processing techniques for estimating
the spatial channel from the signals received at the
base station antenna array. These techniques
conventionally involve an inversion or singular value
decomposition of a signal covariance matrix. The
35 computational complexity of this calculation, however,
is so high that it is presently not practical to

implement. These highly complex approaches capitalize
on the theory of array signal processing. This
approach estimates the uplink channel (e.g., the angles
and times of arrival of the multipath signal parts) to
5   create a space-time matched filter to allow for maximum
signal delivery. The proposed method involves
computation of a signal covariance matrix and
derivation of its eigenvectors to determine the array
coefficients. The basic problem of array signal
10 processing is formulated in the following expression:

X=AS+N

where **X** is a matrix of antenna array signal snapshots
15 (each column incorporates snapshots of all antenna
elements), **S** is the transmitted signal matrix (each
column incorporates snapshots of the information
signal, **A** is the antenna array and channel response or
array manifold matrix, and **N** is the noise matrix. The
20 main challenge of array signal processing is to
estimate **S** based on the statistics of **A** and **S**, that is,
to reliably and correctly estimate all the incoming
signals with the presence of interference and thermal
noise, **N**. This problem has been a subject for
25 extensive research for several years. Two well known
estimating algorithms involve Maximum Likelihood
Sequence Estimation (MLSE) and Minimum Mean Square
Error (MMSE). Using these techniques, if **S** represents
signals with known properties, such as constant modules
30 (CM) or finite alphabet (FA), the process can be
executed using the temporal structure statistics of the
known signal. If the array manifold is known, then
convergence can be made faster. This process, however,
is very computational intensive. In a base station
35 that is required to simultaneously support more than

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100 mobile units, the computation power is presently
beyond practical realization.

    Most adaptive beam forming methods described in
the art (e.g., U.S. Pat. No. 5,434,578) deal

5   extensively with uplink estimation, while requiring
extensive computation resources. Few, however, deal
with downlink estimation, which is a more difficult
problem. Because the spatial channel is frequency
dependent and the uplink and downlink frequencies are

10 often different, the uplink beamforming techniques do
not provide the base station with sufficient
information to derive the downlink spatial channel
information and improve system capacity. One technique
for obtaining downlink channel information is to use

15 feedback from the subscriber. The required feedback
rates, however, make this approach impractical to
implement.

    There is a need, therefore, for increasing
wireless system capacity using beamforming methods that

20 overcome the limitations discussed above in the known
approaches.

SUMMARY OF THE INVENTION

    The present invention provides a method for

25 wireless communication that exploits the spatial domain
in both uplink and downlink without requiring
computationally complex processing. The method
provides for significant capacity enhancement in both
uplink and downlink while maintaining implementation

30 simplicity. This goal is achieved by eliminating the
necessity for covariance matrix processing, using low
bit count arithmetic and by capitalizing on signal
multipath structures.

    A method for wireless communication according to

35 the present invention comprises transmitting from a
mobile unit a code modulated signal, such as a CDMA

signal, which is obtained by modulating original
symbols by a predetermined pseudo-noise sequence. The
original symbols represent an original information
signal. A base station antenna array then receives in
5   parallel N complex-valued signal sequences from N
corresponding antenna elements. Each of the N signal
sequences are then correlated with the pseudo-noise
sequence to de-spread and select N received signals
comprising N received symbols corresponding to a common
10   one of the original symbols. The N received symbols
are then transformed in parallel to obtain N complex-
valued transformer outputs which are then correlated
collectively with a set of complex array calibration
vectors to obtain spatial information about the signal.
15   Each array calibration vector represents a response of
the antenna array to a calibration signal originating
in a predetermined direction relative to the base
station. The above steps are repeated to obtain
spatial information (angle of arrival (AOA), time of
20   arrival (TOA), and distance to the mobile unit) about
multiple signal components corresponding to the same
mobile unit. This spatial information is then used to
spatially filter subsequent complex-valued signal
sequences. The filtered signal is then demodulated to
25   obtain a symbol from the original information signal.
        The original symbols are selected from a finite
symbol alphabet. In a preferred embodiment, the finite
alphabet contains not more than 64 symbols and the
calibration vectors comprise complex-valued components
30   having 1 or 2-bit-plus-sign real part and 1 or 2-bit-
plus-sign imaginary part. The number of bits can be
increased if necessary. This simple representation
allows computing the correlation via addition without
the need for computationally complex multiplications.
35   In one embodiment, the correlating step yields spatial
information about multiple signal components from the

mobile unit having small time separated signal parts
(i.e., having a time spread less than one chip).
Another embodiment of the invention includes the step
of tracking time and angle information of the multiple
5   signal components. In one embodiment, an analog signal
with known amplitude and zero phase (i.e., a "sounding"
signal) is inserted into transmission and receive
channels of the base station. The signal at each of
the channel outputs is then decoded to determine its
10 phase and amplitude. The measured phase and amplitude
data are then used to correct the antenna calibration
data (array manifold matrix), thereby eliminating phase
and amplitude mismatch in a multi-channel receiving and
transmitting system, which can be due in part to
15 temperature changes, component degradation, receive and
transmit power, etc.
     The invention further provides for a spatially
filtered downlink information signal in accordance with
the spatial information about the multiple signal
20 components that was determined from the uplink. The
spatial filtering comprises assigning the mobile unit
to a beam based on spatial information about the mobile
unit. This spatial information comprises directional
and distance information about the mobile unit. The
25 downlink beams are a dynamically adaptive set of
overlapping broad and narrow beams such that closer
mobile units are assigned to broader beams and more
distant mobile units are assigned to narrower beams.
The downlink beam width is determined based on uplink
30 signal AOA distribution (collecting many symbols) and
if possible, distance. Since normally AOA spread is
associated with distance such that the farther the
mobile unit the less AOA spread, the AOA spread can be
used as mentioned above. The set of beams are modified
35 depending on the statistics of the spatial information
of all mobile units served by the base station in order

to optimize system performance. In one embodiment,
multiple (2 to 4) narrow beams (2 to 3 degrees) are
formed in a wide aperture antenna array to cover a
scattering zone to minimize the effects of fast

5   (Rayleigh) fading. The wide aperture array allows
multiple narrow beams in the same general direction to
be constructed with low correlation weight vectors,
providing low correlation (approximately 0.7 or less)
between the beam outputs within the wide antenna array.

10 The width and orientation of the beams are determined
by evaluating the angular spread of the incoming
signal, and in particular, by determining the peak and
spread or variance of an angle of arrival histogram.
In the preferred embodiments, the transmitting of the

15 downlink beams is performed in accordance with
beamforming information comprising complex-valued
elements having 3-bit-plus-sign real part and 3-bit-
plus-sign imaginary part. In the preferred embodiment
for CDMA IS-95, downlink traffic beam is assigned to

20 specific mobile units while the overhead beams are
maintained as for three or six sector base stations. A
small phase is maintained between traffic and pilot
beams to prevent degradation of demodulation
performance at the mobile station.

25   In some embodiments, a pilot signal is code
multiplexed into the signal transmitted from a mobile
unit to the antenna array of the base station for wide-
band CDMA communication systems. The base station
correlates the pilot signal of the incoming signal with

30 a sequence of delayed pilot signals generated from the
base station. These correlation values are spatially
correlated with the antenna array manifold matrix to
produce a signal angle of arrival (AOA) and time of
arrival (TOA) histogram. The resulting histogram is

35 used to determine the "best" AOA and TOA for forming
uplink and downlink beams directed to the desired

scattering zone. Using a pilot signal, instead of the
actual signal itself, for spatial correlation, results
in a simpler communication system. In many situations,
the AOA histogram can be compiled from a spatial
5   correlation between the array response vector
(comprising electrical amplitude and phase of all
antenna array elements) and the array manifold matrix.
Knowledge of the array manifold is useful due to the
relatively small angular spread between signals at
10 longer distances.
      The invention also provides a CDMA base station
implementing the above method. The station comprises
an antenna array having N antenna elements and a set of
N receivers coupled to the N antenna elements to
15 produce N incoming signals. The base station also
comprises a set of N de-spreaders coupled to the N
receivers for producing from the N incoming signals N
de-spread signals corresponding to a single mobile
unit. A set of N symbol transformers is coupled to the
20 N de-spreaders and produces a complex-valued output
from the de-spread signals. A spatial correlator
coupled to the N symbol transformers correlates the
complex-valued output with stored array calibration
data to produce beamforming information for multiple
25 signal parts associated with the mobile unit. In the
preferred embodiment, the array calibration data is
composed of complex-valued array response elements
represented as bit-plus-sign imaginary parts and bit-
plus-sign real parts. A receiving beamformer coupled
30 to the spatial correlator and to the N receivers then
spatially filters the N incoming signals in accordance
with the beamforming information. A RAKE receiver (or
other equivalent receiver) coupled to the receiving
beamformer produces an information signal from the
35 spatially filtered signals. In one embodiment, the
base station also includes a tracker coupled to the

spatial correlator and to the receiving beamformer.
The tracker tracks multiple signal parts and optimizes
the performance of the receiving beamformer.

5   In the preferred embodiment, the base station also
includes a transmitting beamformer coupled to the
spatial correlator. The transmitting beamformer
generates spatial beams in accordance with the beam-
forming information to increase system capacity. The
spatial beams are a dynamically calculated set of
10   downlink beams comprising narrow beams and overlapping
broad beams such that the narrow beams are phase
matched to the overlapping wide beams. The spatial
beams are selected such that more distant mobiles are
assigned to narrower beams and closer mobiles are
15   assigned to broader beams.

In one embodiment, the base station comprises a
compensation signal source and a compensation detector
both coupled between the set of transmitting and
receiving beamformers and the set of N transmitters and
20   receivers. The compensation signal source injects a
known amplitude, zero phase analog "sounding" signal
into the transmission channels while the compensation
detector decodes the sounding signal and accumulates
measured phase and amplitude data, which are used to
25   correct phase and amplitude mismatch data.

DETAILED DESCRIPTION

Although the following detailed description
contains many specifics for the purposes of
30   illustration, anyone of ordinary skill in the art will
appreciate that many variations and alterations to the
following details are within the scope of the
invention. Accordingly, the following preferred
embodiment of the invention is set forth without any
35   loss of generality to, and without imposing limitations
upon, the claimed invention.

Fig. 1 provides a general view of the system
architecture of a base station according to the present
invention. The base station comprises a receiving
antenna array 10 having N antenna elements. In this
5   embodiment, the system also comprises a separate
antenna array 15 for transmission. Using antenna
duplexers, however, the arrays can be combined, as is
well known in the art. The embodiment allows for low
cost duplexers and antenna filters since much less
10 power is required per element to provide the required
effective radiated power (ERP) due to beam-forming.
Preferably, the number N of antenna elements is
approximately 16.

Each of the N antenna elements is coupled to a
15 corresponding one of a set of N conventional receivers
101. Each receiver down-converts an incoming signal in
frequency and digitizes the signal to produce a
received signal having I and Q (in phase and
quadrature) signal components. In this embodiment, the
20 receivers are coherently tuned by a common local
oscillator 104 to allow for both phase and amplitude
data measurement, thereby producing, at any given
instant, an N-dimensional received signal vector having
complex-valued components. Alternatively, a
25 calibration signal of fixed frequency can be injected
to all receiver channels simultaneously with the
received signal, allowing for continuous estimation of
the phase and amplitude difference among the receivers.
The calibration signal can be differentiated from the
30 received signal since it is not spread and can have a
very low level since its integration can be very long.
Specific relevant receiver designs are presented in
U.S. Pat. No. 5,309,474, which is incorporated in its
entirety.

35   The received signal vector from the N receivers
101 is fed to a set of L channel estimators 11 and also

to a corresponding set of L receiver banks 14. Each
channel estimator 11 and corresponding receiver bank 14
is used to estimate the channel and receive the signal
from a single mobile unit. Thus the maximum number of
5   mobile units that can be simultaneously served by the
base station is L. In a preferred embodiment, L is at
least 100. The estimators 11 are identical to each
other in both structure and principles of operation.
Similarly, the receiving banks 14 are also identical.
10   Accordingly, the following description is limited to a
single estimator 11 and its corresponding receiver bank
14 which serve to estimate the channel of a single
mobile unit and receive its signal.

    In the preferred embodiment, channel estimator 11
15   comprises a set of N de-spreaders 102, a corresponding
set of N fast Hadamard transformers (FHTs) and a
spatial correlator 105. The de-spreaders 102 are
conventional code-correlators described in detail, for
example, in U.S. Pat. No. 5,309,474. Each of the N de-
20   spreaders correlates a single component of the received
signal vector with a pseudo-noise (PN) code sequence
assigned to the associated mobile unit in accordance
with the IS-95 CDMA standard. Each code correlator or
desperado 102 uses a variable time offset (synchronized
25   with the other code correlators in the same bank) to
separate multipath parts that arrive with at least one
PN chip period difference. The time offset is
determined by repetitive hypothesis, e.g., setting the
code time offset and collecting the symbol length of
30   the samples, then executing the process described
herein. The result is the CIR buffer (described
later), where the peaks represent the different signal
paths' TOAs. The following description discusses the
processing of one multipath part. All multipath parts
35   that are strong enough to be isolated are processed
identically.

                        /5

Each desperado 102 outputs a de-spread signal
corresponding to one mobile received at one antenna.
This de-spread signal is fed into a fast Hadamard
transformer (FHT) 103. The FHT used in the present

5   invention is identical to conventional FHTs (described,
for example, in U.S. Pat. No. 5,309,474), except that
the FHT of the present invention retains the complex
phase information of the input. In other words,
whereas the standard FHT outputs are converted to

10   magnitudes, the FHT used in the present invention
outputs complex numbers, thereby preserving both phase
and amplitude data. Each FHT in this embodiment has 64
complex outputs, whose magnitudes represent the degree
to which the de-spread signal correlates to each of the

15   64 symbols in a predetermined symbol alphabet. In the
preferred embodiment, the symbol alphabet is a set of
64 orthogonal Walsh symbols.

For a given symbol received at the antenna array
10 (in IS-95, a symbol period is approximately 208

20   microseconds), the signals received at the N antenna
elements are, separately and in parallel, passed
through N respective receivers 101, de-spreaders 102,
and FHTs 103, while retaining the relative phase
information of the signals. The collection of N FHTs

25   103 together produce an N x 64 signal matrix B of
complex elements. Each column of B is an N-dimensional
vector, called the spatial response vector, whose N
components represent the correlation of one Walsh
symbol with the signal received at the N antenna

30   elements. The matrix B is fed column-by-column to the
spatial correlator 105 following timing synchronized to
the Walsh symbols.

As will be described in detail below in reference
to Fig. 4, spatial correlator 105 correlates the signal

35   matrix B with an array calibration matrix A. The
matrix A is obtained off-line by calibrating the

antenna array for phase and amplitude vs. angle. The
correlation produces a correlation matrix **C** that
represents the correlations of the signal received at
the antenna array with both a set of predetermined

5   directions and a set of predetermined symbols. From an
analysis of the matrix **C**, the correlator 105 produces a
signal angle of arrival (AOA) and a scalar value (AOA
quality) that is proportional to the "purity" of the
wave front and the signal level. This data is

10   transferred to a controller 106 that uses it to
determine the best uplink beam coefficients for this
particular signal part. Typically, this entire process
is performed for the four strongest multipath parts by
setting the code and the "start of sampling period"

15   time to the expected TOA, as is known in the art. In
addition, time of arrival (TOA) and AOA certainty data
are produced, allowing for the generation of a
spatially matched filter that contains beam-forming
information for each signal part. AOA results are

20   collected by repeating the above process for many
incoming information symbols. This data is used to
produce an AOA histogram from which the most expected
AOA and AOA distributions are calculated for every
individual signal part. The AOA provides beam

25   direction information, and the AOA distribution
provides beam width information. The functions of the
channel estimator 11 described above are performed in
parallel with all the other channel estimators for the
other mobile units being handled by the base station.

30   The controller 106 receives beam-forming
information from each of the channel estimators 11.
Thus controller 106 obtains spatial information
regarding all the signal parts from all the mobile
units. The controller 106 then downloads this

35   information, in the form of coefficients, to the
receiving banks 14 which use the spatial information

from the channel estimators 11 to improve the reception
of the signals from the mobile units. Each receiving
bank 14 comprises beamformers 112 to form narrow beams
towards the signal parts associated with a single
5   mobile unit. Because the strong signal parts are
selectively detected, the beamformer creates a well-
matched spatial filter for the incoming signal,
including its multipath components. The beamformers
112 feed spatially filtered signals to the four fingers
10   of a conventional IS-95 RAKE receiver 113 (described in
U.S. Pat. No. 5,309,474). It should be noted, however,
that the beamformer outputs can be fed to other
receiver types, which are known to those skilled in the
art. As a result of the spatial filtering process
15   described above, the carrier-to-interference (C/I)
ratio is significantly improved over conventional CDMA
systems. The improvement in C/I is about the ratio
between the effective beamwidth created (about 10 to 30
degrees) to the existing antenna beams (about 100 to
20   120 degrees). Note that the AOA and TOA data are also
transferred to a central controller 120 where the
system determines the most optimal downlink beam
configurations. The downlink process will be discussed
later, as part of the description of Fig. 4.
25   In another embodiment, controller 106 can assign
narrow beams (typically 2 to 3 degrees in width) within
the wide aperture antenna array to cover different
sections of a scattering zone for fading mitigation.
The typical scattering zone around a mobile transmitter
30   is described by a circle with about 30 to 100
wavelength size radius. Large reflectors in the
neighborhood (such as very large buildings or
mountains) can create secondary scattering zones that
produce time differentiable (by spread spectrum
35   receiver) multipath propagation, and hence, provide for
multiple scattering zones. In traditional spatial

diversity, the signal is collected at different points
in space where the arriving multipath is combined with
different phases. Thus, when one antenna exhibits a
destructive combining situation, the other antenna has
5   a high probability of exhibiting a desirable
constructive combining situation.

By forming beams that are narrow enough to
distinguish a collection of energy emanating from
different sources, while still generally pointing the
10   beams in the same direction, fading from multipath
propagation can be greatly reduced. With a wide
aperture array, the array weight coefficients can be
changed for different beamformers to alter the beams
within the array. Even though narrower beams are
15   formed in the antenna array, high grading lobes are not
considered a significant problem in the case of CDMA
since the interference is the summation of all other
active subscriber energies, and since the grading lobes
are narrow by nature, they are mostly rejected.
20   If the beamwidth is narrow enough relative to the
scattering zone size, different populations of
multipath sources participate in each beam, and hence,
the power/time function at each beam will not be
correlated with the other beams. Scattering zones are
25   typically between 5 and 10 degrees in angle, but can
vary in size, depending on factors such as the distance
from the base station and area characteristics. For a
5 to 10 degree scattering zone, a 3 to 6 degree
beamwidth allows adequate distinction between other
30   beams. Since the typical RAKE receiver can accept four
antenna beams, this embodiment provides for
simultaneous non-correlated power/time function
processing. Simulation results, such as shown in Figs.
2 and 3, show that the effectiveness of this method is
35   very similar to the current spatial diversity method.

For small angular spread conditions (e.g., 2 to 3
degrees) associated with a small scattering zone or
distant subscriber, two to four narrow beams (e.g., 3
to 6 degrees in width) are contiguously arranged with
5   some overlap to cover the scattering zone (e.g., 5 to
10 degrees in angle). Fig. 2 shows the cumulative
probability density function (PDF) or cumulative
distribution function (CDF) for this arrangement, which
represents the power distribution of symbol amplitude
10   as received by a given system. The four curves
represent the CDFs for various systems: the solid
curve on the left 50 is for a system using standard
spatial diversity, the dashed curve 51 represents a
system using a single beam directed at the center of
15   the scattering zone, the dashed-dotted curve 52
represents a system using a single beam quickly
tracking the changing AOA associated with the varying
multipath, and the solid curve on the right 53
represents a system using the multi-beam arrangement of
20   the present embodiment. As seen from the shapes of the
CDF curves, the effectiveness of the multi-beam
arrangement is similar to that of a single tracking
beam, which requires very high processing power, and
standard spatial diversity. Note that the horizontal
25   axis represents relative gain, which can be changed by
using different types and/or numbers of antenna
elements, etc.
        As the angle spread increases, the angular
separation between the beams increases, as do the beam
30   widths. However, the number of beams remains the same.
The beams are angularly spread to sample different
sections of the scattering zone. In addition, the beam
width is increased, limited by the size of the antenna
array. As Fig. 3 shows, which represents the CDFs
35   using the same systems as in Fig. 2, but with a wider
angular spread (10 degrees), the effectiveness of the

fading mitigation using the multi-beam arrangement
increases when the angular spread increases (due mostly
to smaller.distances between the mobile unit and base
station). As the angular spread increases,
5   corresponding to an increase in the scattering zone
viewing angle, the correlation between beam outputs
decreases since each beam can be pointed to cover areas
more widely spaced apart. As a result, there is a
lower correlation between beam outputs, which improves
10   on diversity efficiency or fading mitigation, as
discussed in "Mobile Cellular Telecommunications" by
William C.Y. Lee, incorporated by reference in its
entirety.

The angle/time of arrival estimation described
15   above and in additional detail below allows for both
single scattering zone and multi-scattering zone
handling. Angular spread can be determined in real
time by way of histogram processing of angle of arrival
samples. When fading is produced by a large scattering
20   zone, the angle of arrival results (AOA samples) are
distributed with large variation (and can be estimated
by the variance of AOA results). The main AOA,
however, can be estimated with the histogram center of
gravity. The histogram center of gravity is determined
25   by "smoothing" the histogram through a low pass filter
(e.g., Hamming, Raised Cosine, etc.) and finding the
maximum point of the "smoothed" histogram.

Thus, in the above described embodiment, an
arrangement of multiple narrow beams in a wide aperture
30   array provides fading mitigation by multiple beams for
a highly directional antenna array since moderate
beamwidths (e.g., 10 degrees or more) will not provide
diversity because the beams "encapsulate" the whole
scattering zone and hence cannot provide for non-
35   correlated multipath combining.

Although the preferred embodiment uses an IS-95
based architecture, the above process can be
implemented with any wireless protocol that makes use
of a finite alphabet or training sequence. For

5   example, in GSM systems a training sequence is
available in every wireless burst. Since the training
sequence is known, a correlation between the incoming
signal and a stored training sequence at the receiver
will produce the same results as described above

10 (provided frequency error is not too great relative to
the sequence length). The desperado 102 and the FHT
103 are replaced in this case by a training sequence
correlator (convolver). Since there is only one
possibility for a training sequence, it is not required

15 to try for many possibilities as done by the Hadamard
transformer in the preferred embodiment. Systems with
training sequences for use with the present invention
are discussed in more detail in a later portion of the
description.

20   Fig. 4 illustrates the details of the spatial
correlator 105. In this embodiment the spatial
correlator is a stand-alone unit. Due to redundant
functionality between this unit and the current
implementation of IS-95 RAKE receiver, however, the

25 spatial correlator can be integrated with the RAKE
receiver. The preferred embodiment is optimized to IS-
95 (uplink utilizing M-ary modulation), however, any
signal that has either a finite alphabet (limited
number of symbols) or a training sequence can utilize

30 the same idea. The use of the known signal structure
facilitates simple array response vector determination
and eliminates the necessity for complex covariance
matrix calculation and analysis. Hence, this approach
can be utilized for GSM and TDMA wireless air-

35 interfaces as well.

The columns of the signal matrix **B** (i.e., the
spatial response vectors from the FHTs) pass through a
multiplexer (MUX) 206 and are then correlated with the
columns of the array calibration matrix **A** which is
5 stored in a random access memory (RAM) 203. In
abstract terms, the correlation process is performed by
multiplying the conjugate-transpose (Hermitian) of the
calibration or array manifold matrix **A** by the signal
matrix **B**. The result is a correlation matrix $C = A^H B$.
10 It is important to note that this abstract calculation
may be implemented in many different ways, all of which
are mathematically equivalent to each other. The
calibration matrix **A** is also known as the array
manifold matrix and is generated by measuring the
15 antenna array response in an antenna test range. Each
column of **A** represents the response of the antenna
array in one of a predetermined set of directions. For
example, if the angular space is divided into 360
directions, then each of the 360 columns of **A** is an N-
20 dimensional vector representing the response of the N
antenna array elements in one of 360 directions from
the array. In the computation of the matrix **C**, these
360 vectors are spatially correlated with the 64
columns of the signal matrix **B** to produce a 360 x 64
25 element matrix, where element i•j represents the
correlation of the received signal with the jth symbol
in the ith angular direction.

        In the preferred embodiment, the correlation is
performed very efficiently through the use of a unique
30 and simple calibration table representation which
allows the matrix multiplication to be implemented
without any multiplications. Each complex-valued entry
of the calibration table matrix **A** is quantized such
that both real and imaginary parts are each represented
35 by two bits only. More specifically, each part is
represented by two bits, one numeric bit and one sign

bit, thus: (0,0)=-0, (0,1)=+0, (1,0)=-1, (1,1)=+1.
Each complex-valued entry is therefore represented by
just four bits. The reduced resolution in this simple
quantization scheme is compensated by increasing the
5   number of array elements to about twice relative to
current base station arrays. This simple bit-plus-sign
data structure allows the vector dot products between
the matrix columns to be calculated using a complex
adder 204. In conventional implementations, the vector
10   dot product would require a collection of N
multipliers. The technique of the present invention,
therefore, dramatically simplifies the implementation
of the spatial correlation operation.

The complex-valued entries of the calibration or
15   array manifold matrix A may be subject to errors due to
the analog parts of the transmitting and receiving
channels experiencing non-predictable changes from
factors such as temperature change, system component
degradation, variations in transmit and receive power,
20   etc. By measuring the phase and amplitude response of
the channels, the behavior of the receiving and
transmitting channels can be known, thereby allowing
correction of the entries of matrix A, which represent
the response of the N antenna array elements in a given
25   direction from the array. Measurement of the phase and
amplitude response of a signal channel requires a
"sounding" operation, i.e., injection of an analog
signal into the channel (with characteristics that are
matched to the channel frequency and amplitude
30   response) and determination of signal amplitude and
phase at the channel output.

In the case of an analog or TDMA base station,
injection of a sounding signal may interfere with the
on-going data transmission. If the sounding signal is
35   made low, the sounding accuracy will degrade. CDMA
communication allows for "embedding" the sounding

signal within the general data flow without losing
sounding accuracy or interfering with the main data
signal. Since the data signal is coded-spreaded (IS-95
or similar), the sounding signal can be either non-
5   modulated or coded-spreaded, with statistical
orthogonality to the data signal. A "matched
accumulator" (using a matched de-spreading code) on the
channel output allows for coherent decoding of the
sounding signal (to determine its phase and amplitude),
10 while the data signal contribution to the detector
output (being randomly distributed in phase and
amplitude) is nullified. The measured phase and
amplitude data can be used to correct the analog
channel response, thereby eliminating phase and
15 amplitude mismatch in a multi-channel receiving and
transmitting system.

In one embodiment of this method, shown in Fig. 5,
compensation circuits 501 and 502 are coupled between
the N transmitters 109 and the transmitting bank 12
20 (Fig. 1) and between the N receivers 101 and the
channel estimators 11 (Fig. 1), respectively.
Compensation signal source circuit 501 provides
sounding signals into the transmit (TX) and receive
(RX) channels. A constant generator A 503 in
25 compensation signal source circuit 501 supplies a
constant value A to a test transmitter 504, providing
test transmitter 504 with a signal of a known amplitude
and a zero phase. A constant generator B 505 in
compensation signal source circuit 501 provides a
30 constant value B to selected ones of transmitters 109
for channel response evaluation.

For compensation in the receiving channels, the
output of test transmitter 504 is frequency converted
to match the RX modules using a frequency converter
35 module (FCM) 506, i.e., eliminating the phase and
amplitude differences between the transmitter and RX

modules. This can be done by measuring these values
and then compensating for them during the matrix
calculations. FCM 506 injects the sounding signal to
all receiving channels through an equal phase and
5   amplitude power divider 507. Each of a set of N
couplers 508 couples the sounding signal with
corresponding antenna elements from receiving antenna
array 10. The signals from each of the N couplers 508
is then fed to the associated one of the N receivers
10   101 for down-conversion to produce digital signals
having I and Q signal components. The set of N signals
can then be placed on a receiving bus for use, such as
inputs to channel estimators 11 and receiving banks 14
of Fig. 1.

15   The received output or channel under evaluation
(digital output) is selected and multiplied by a signal
generated from a constant generator A' 509. The signal
from constant generator A' 509 is made equal to
constant generator A 503 to decode or de-spread the
20   sounding signal from the received data signal. Then,
the digital values (I and Q) are accumulated by a
compensation detector accumulator 510. The
accumulation process period is limited only by the
channel response variation rate (assumed to be very
25   low) and the size of the registers in the accumulator
510. Hence, the accumulation process provides an
integration time sufficient to extract the sounding
signal out of the signal mixture on the receiving
channel under evaluation. The sounding signal "can be
30   -30 dB relative to the total signal energy in the
channel. By coherently decoding the sounding signal,
the phase and amplitude of the sounding signal can be
measured to determine the phase and amplitude response
of that particular receiving channel. For example, I
35   and Q samples of the RX are directly accumulated for a
determined integration period. The I and Q data

includes both amplitude and phase of the measured
channel.

    The procedure described above is repeated for each
receiving channel until the phase and amplitude
5   responses of all the receiving channels are known. The
channel compensation responses for each channel are
then combined to form a "compensation vector", which
can be used to correct the amplitude and phase of the
measured or manipulated data. Entries in the
10 calibration matrix **A**, stored in RAM 203 (Fig. 4), can
be corrected by dividing each row of matrix **A** by the
corresponding row vector (complex value) of the
compensation vector. This operation results in
corrected calibration matrix, eliminating errors in all
15 the receiving channels.

    A similar procedure is utilized for compensation
in the transmitting channels. A transmission channel
selector 511 in compensation signal source circuit 501
selects a transmitting (TX) channel from a transmitting
20 bus, which can emanate from transmitting banks 12 (Fig.
1). A constant (which can be very small) from constant
generator B 505 is added to the signal from the
selected channel to be compensated. The constant
signal is alternated between positive and negative
25 values such that constant generator B 505 is equal to a
constant generator B' 512. The resulting signals are
then converted by the set of N transmitters 109, sent
to a set of N couplers 508, and combined at a power
combiner 513. The combined signal is frequency
30 converted by FCM 506 and then down-converted by a test
receiver 514 to produce digital I and Q signal
components. The combining, although it may degrade the
SNR conditions, allows for a total passive arrangement
that is very important when the antenna array is
35 located at the top of a tower (which is often not
easily accessible).

The output of test receiver 514 is multiplied by a
signal generated from a constant generator B' 512. The
signal from constant generator B' 509 is made equal to
constant generator B 505 to decode or de-spread the
5   sounding signal from the transmission channel. Similar
to correlation of the receiving channels described
above, phase and amplitude responses from the
transmission channels can be used to correct
transmission coefficients from a transmission
10   coefficient table (not shown). Thus, the compensation
system of Fig. 5 evaluates channel responses of both
transmit (TX) and receive (RX) sections of the base
station.

        Referring back to Fig. 4, timing generator 201
15   synchronizes the spatial correlator process to the
Walsh symbol period (i.e. the end of the Hadamard
transform) that is derived from the base station pilot
timing. The N x 64 signal matrix is latched into a MUX
circuit 206 which provides the column vectors to the
20   complex adder 204 one at a time. For each vector, the
complex adder 204 performs separate correlations of the
vector to every one of the columns of the calibration
matrix A. Because the calibration matrix data are only
0, 1, or -1, the data are used in the complex adder 204
25   to decide whether to respectively null, add, or
subtract each element in the row vector. A RAM address
generator 202 is driven also by the same timing
generator 201 to synchronize the presentation of the
columns of calibration data with each latched vector.
30   Note that the number of array elements N does not
change the correlation matrix dimensions, which are
determined only by the number of predetermined symbols
in the alphabet and the number of pre-defined angular
directions. The correlation matrix C is stored in a
35   spatial correlation RAM 207 and processed by a maximum
value selector 205 that is a simple serial comparator

in the preferred embodiment. The end result of the
spatial correlator process is the best expected AOA for
the selected signal part and an associated "inner
product" value (used as a certainty factor). This
5   result is reported to the controller 106 (Fig. 1) only
if a preset threshold, discussed below, has been
crossed. This threshold value is updated from time to
time as necessary. When the threshold has been
crossed, the controller registers the time offset
10   associated as the signal part TOA. This information is
used to estimate the mobile unit range from the base
station. It is possible to identify more than one
maximum at a time utilizing a recursive process: after
identifying the largest value in the correlation
15   matrix, the neighboring matrix elements are ignored
(ignoring the neighboring elements minimizes the
probability for "non peak" selection) and another
"peak" search is executed. This feature allows the
identification of multipath parts that cannot be
20   differentiated in time alone (as done in existing RAKE
receivers), allowing for beam-forming reception of
small time spreaded multipath. This approach has great
advantage for close to base station mobile unit
communication.
25   The threshold value is calculated by averaging the
reported results, I and Q, over a long averaging period
"window". For example, K reported results are
accumulated at the controller 106, and the accumulated
result is divided by K. Since most of the reported
30   results are generated by non-time correlated elements,
the results are "noise-like", and averaging them
provides a good estimate of the channel noise level.
Since the channel noise is a linear function of the
number of active mobile units, this level needs to be
35   updated from time to time as stated.

Fig. 6 details the uplink beamformer 112 of Fig.
1. In this embodiment, the uplink beamformer is
presented as a stand-alone unit. However, it is
possible to integrate the beamformer 112 into the
5   channel estimator 11 due to the "bit-plus-sign"
arithmetic that makes it a very low gate count device.
Signal outputs from the N base station receivers 101
are fed into a complex adder 604 for beam-forming.
Since the data rate for IS-95 is about 10 Mega-samples
10   per second, the complex adder 604 can execute at least
four vector sums per one vector data sample using
present technology. The beam-forming coefficients are
downloaded from the controller as described above into
a coefficient RAM 603. A timing generator 601 and an
15   address generator 602 cause the coefficients to
"rotate" into the complex adder 604. The coefficients
are used as described above, in reference to the
spatial correlator of Fig. 4, to form a dot product
using only complex addition. The vector summation
20   result is fed into an interface unit 605 for
transferring the result to the RAKE receiver modem. In
other embodiments, any finite alphabet or training
sequence protocol based modem could be used. The
effect of the beamformer 112 is to spatially filter the
25   incoming signal to preferentially select for signals
arriving from the known directions of the signal parts
of a particular mobile unit. Signals from other
directions are attenuated, and the reception of the
desired signal is improved.
30   Fig. 7 illustrates an example of a spatial
distribution of downlink beams. Downlink management is
quite different from the uplink since IS-95 is not a
symmetrical protocol and uplink frequency is different
from the downlink frequency by at least 60 MHz
35   (cellular). The difference in frequency causes the
uplink and downlink channels to be non-correlated. The

AOA and TOA of the uplink and downlink, although
statistically similar, may differ significantly.
Hence, the downlink can be only statistically estimated
based on data collected in the uplink, as described
5   above.   In addition, the downlink requires broadcasting
of a pilot signal to associated mobile units.   As a
result, individual downlink beams are not possible;
only "mobile group" beams are realizable.   Hence, the
downlink approach is based on a combination of wide and
10   narrow beams determined by data collected in the
uplink.

        Thus, referring to Fig. 1, transmitting
beamformers 117 in transmitting banks 12 are coupled to
the spatial correlator 105 via the central controller
15   120 to receive AOA and TOA data for generating spatial
beams in accordance with the beam-forming information.
The spatial beams are selected from a set of calculated
beams comprising narrow beams and overlapping broad
beams, where the narrow beams are phase matched to the
20   overlapping wide beams.   The beamformers 117, which are
conventional digital beamformers, take signal samples
(scalar I and Q) and multiply them by the weight vector
to produce a vector, where each element includes a
scalar represention of the signal going into the
25   individual antenna.   Routing circuits 116 and routing
and summation circuits 115 are data switches to route
signals coming from the plurality of transmissions into
the beamformers 117 and the transmitters 109.   The beam
configuration is determined by the mobile unit
30   distribution around the base station.   The wide beams
are required to assure proper coverage at close
proximity to the base station where most of the
downlink signal arrives to the mobile unit by way of
scattering from nearby reflectors.   The system adjusts
35   the wide beams 701 to assure proper coverage for the
mobile units close to the base station.   The narrow

beams 702 are adjusted mainly to accommodate "far away"
mobile units. Since most of the mobile units will be
in the outer coverage area, the narrow beams are
expected to service the majority of the mobile units.

5   Increasing the number of downlink beams causes the
increase of softer hand-off, thereby countering the
increase in capacity. Hence, assigning beams in the
downlink must be done very carefully.

The increase in downlink capacity can be estimated
10   as follows:

$$Q*P+Q*P/N+X*P+X*P/N+(Q(1-P)/N+X(1-P)/N)*(1+B) = Q$$

15   $$\Psi = \frac{Q+X}{Q} = \frac{1}{P+P/N+(1-P)(1+B)/N}$$

We assume a uniform mobile unit distribution, and a
maximum illumination of Q, that is the maximum number
of simultaneous transmission channels including softer
20   hand-off.

The term $Q*P$ is the number of mobile units that
come in with high angular spread, called "Wide Angles".
$Q*P/N$ is the portion of the portion of Wide Angles that
are within the narrow beam, and are all in softer hand-
25   off, thus, adding to the illumination in the overlapped
sector twice.

If, as a result of the beams combination X mobile
units can be added, $X*P$ additional Wide Angle types are
added (assuming P remains as before), out of, $X*P/N$ are
30   following the same rule as for the $Q*P/N$ above.

In the narrow beam space, we get $Q(1-P)/N+X(1-P)/N$
mobile units, but due to some hand-off caused by the
overlap we must increase the value of their
illumination by factor 1+B. B is the ratio of the
35   number of users in handoff to the total number of

32

users, which is determined experimentally. B can be
kept small since outer cell associated mobile units
will naturally prefer the narrow beams.

        Fig. 8 is a graph of the capacity increase ratio
5   with respect to both the number of narrow beams and the
probability/10 of wide angular spread multipath with
softer hand-off probability fixed at 20%. Fig. 9 is a
graph of the capacity increase ratio as a function of
hand-off probability/10 and the wide angular spread
10 multipath probability/10 for four narrow beams. Fig.
10 is a graph of two cases of wide angular spread
multipath for variable hand-off ratio and four narrow
beams. Figs. 8-10 show the improvement in capacity
relative to non-adaptive array base stations.

15   Following the above analysis, the enhancement of
capacity for four narrow beams within one wide beam is
approximately two. If the mobile unit distribution is
non-uniform, the enhancement can be even higher. Fig.
11 is a graph of the expected capacity ratio for
20 different density variance of mobile units. This
improvement requires the narrow beam borders to be
adjusted to avoid mobile unit density peaks. The
borders can be adjusted, for example, according to the
description accompanying Fig. 12 below. This
25 adjustment function must be gradual to avoid excessive
hand-off while changing the beams.

        Fig. 12 presents a flowchart of the downlink beam-
forming determination process. Mobile unit spatial
data is collected and stored in memory in block 1200.
30 This data is then used to evaluate the mobile unit
distribution around the base station by sorting the
data into a two dimensional histogram in block 1205.
The histogram "peaks" are identified in block 1210 as
follows: a two dimensional "smoothing" filter is
35 executed to eliminate noisy histogram "spikes" and a
common two dimensional "peaks search" process is

utilized. For a system that is capable of forming M
downlink beams, M "peaks" are sorted in block 1210.
After the mobile units are sorted by associated pilot
signals in block 1215, the number of mobile units

5   around the.M highest histogram peaks are counted in
block 1220. The mobile unit count is compared to the
closest beam's pilot count for each of the M peaks in
block 1225. Then in block 1230, the mobile unit peak
count is compared to the pilot mobile unit count. If

10   the mobile unit peak count is close to the pilot mobile
unit count, then the next set of spatial information is
stored in block 1200. However, if the mobile unit peak
count is not close to the pilot mobile unit count, then
the closet beam is shifted towards the peak in block

15   1235. The pilot count of the shifted beam is then
compared to the other mobile unit counts in block 1225.
Thus, a closed loop process adjusts the boundaries of
the downlink beams and equalizes the number of mobile
units associated with them. Narrowing the beams will

20   cause some·mobile units to hand-off to different pilots
leaving only mobile units close to the selected "peak"
hanging on to the associated pilot. This process
proceeds at a very slow pace to avoid excessive hand-
off.

25   Fig. 13 presents an apparatus to generate the
antenna array manifold (or calibration) matrix **A**. An
antenna array 1301, incorporating a collection of
antenna elements, is installed on a support mast that
is connected to a turn-table 1304. A controller 1306

30   commands the turn-table to rotate in pre-determined
angle steps or the number of angular directions f the
array manifold **A**. A network analyzer 1305 transmits
through a transmitting antenna 1302 an RF signal with a
particular angle, which is received by the antenna

35   array 1301: The signals received at the elements in
the antenna array 1301 are routed through an RF switch

1303 to the network analyzer 1305 for measurement,
which is well known in the art. In the preferred
embodiment, the antenna array is circular, but the
invention can be implemented with any arbitrary array

5   shape. The RF signal collected for each antenna
element in this case can be written as follows:

$$A_{k,\theta} = F_{k,\theta} \cdot e^{j \cdot \frac{2\pi(R\cos(2\pi k/M - \theta)}{\lambda}}, \text{ where}$$

10   A represents the array manifold function, k is the
element number, θ is the relative angle of arrival
(created by rotating the array relative to the RF
signal source), M is the total number of antenna
elements in the circular array, and λ is the RF signal

15   wavelength. The data is collected and stored in the
controller 1306, which also includes a data storage
unit.

Array manifold information can be used to more
accurately determine multipath angle of arrival (AOA)

20   values and coefficients through spatial correlation in
fast fading environments with high angular spread and
non-predictable multipath. As previously discussed,
spatial processing includes estimating an array
response vector (comprising electrical amplitude and

25   phase of all antenna array elements) of an IS-95 based
CDMA signal to determine multipath angle of arrival
(AOA) values and coefficients through spatial
correlation. These coefficients are then used to
optimally combine a plurality of antenna outputs

30   (through a down-conversion to base band). Thus, the
ability to accurately estimate the array response
vector is an important objective in CDMA systems.
However, the estimation accuracy is limited by the
fading rate (Doppler shift caused by a moving mobile

unit), since the time to collect coherent data is
reduced as the fading rate or Doppler rate increases.
This problem becomes more severe as cellular systems
move from the 800 MHz range to the 1900 MHz range or

5   higher, which can increase the fading or Doppler rate
by a factor of two or more.

In addition, with a frequency division duplexing
(FDD) system, the forward link (transmission from the
base station to the mobile unit) and the reverse link

10   (transmission from the mobile unit to the base station)
occupy different carrier frequencies or bands, but
overlap in time. This difference between the forward
and reverse link frequencies reduces the correlation
between fading of the two links, thereby allowing

15   spatial diversity to be used only with the reverse link
and not the forward link, i.e., array response vector
estimation for forward link array coefficients cannot
be accurately determined.

Various methods for array response vector

20   estimation have been proposed, some of which are
characterized by the degree of knowledge of the
temporal and spatial structure of the signal impinging
on the antenna array. Knowledge of the temporal
structure of the signal (which requires a known

25   training sequence, pilot signal, constant envelope,
etc.) leads to algorithms such as MMSE (Minimum Mean
Square Error), CM (Constant Modulo), etc., which are
sometimes called "blind" or "half blind" estimation
techniques. Blind techniques do not use any apriori

30   information on the signal temporal structure and
antenna array manifold, while half blind techniques can
use temporal structure. A major disadvantage of these
blind methods is the long integration time required for
convergence, especially when the number of interference

35   sources is large (typical for CDMA), which reduces the
efficiency of solutions based on nullifying specific

interferers. Furthermore, using a dedicated pilot
signal at the reverse link requires the pilot signal to
be low power in order to minimize capacity loss in the
reverse link. However, a lower power pilot in coherent
5   demodulation requires a longer integration time to
assure sufficient reference signal quality. Also, an
unknown or varying signal time of arrival (TOA)
requires continuous "time searching", and hence,
prohibits a slow convergence process at each time
10 hypothesis. In CDMA type systems, the signal timing
must be recovered before any demodulation can take
place. Hence, a search process is conducted by a
series of hypotheses through which the system is
varying the time of reference correlating sequence and
15 then cross-correlating with the incoming signal (e.g.,
IS-95C or cdma2000). If matched filter hypothesis is
executed (W-CDMA), the sampling point needs adjustment.
The time required by each hypothesis must be short to
allow a quick search (we assume the determination
20 cannot be done before the spatial process since there
might not be a sufficient signal to noise ratio at that
point).

The types of algorithms mentioned above exploit
the statistical properties of the signal; however, they
25 do not exploit any knowledge of the spatial
characteristics of the array (i.e., the array
manifold). Although the array response vector can
significantly deviate from the array manifold in high
angular spread and non-predictable multipath structured
30 environments, even partial knowledge of the array
manifold could dramatically reduce the required data
integration time and speed the computational process.
Array manifold information allows true data processing
in the spatial domain by exploiting the knowledge of
35 relationships among different antenna outputs to

facilitate simultaneous two-dimensional averaging in time and space.

As mentioned above, in frequency division duplexing (FDD) systems, there is only a statistical
5   connection between the coefficients in the reverse link array response vector and the ideal array response coefficients in the forward link due to the frequency difference between the two links. Hence, determining the forward link transmission coefficients must be done
10   either through an estimation process in conjunction with received power indication feedback at the mobile unit. A power feedback method without initial forward link estimation may be very slow or may not converge at all due to the variable shadowing and fast fading
15   conditions typical of mobile environments.

However, knowledge of the array manifold can be efficiently used to expedite determination of weight vector coefficients for the purpose of signal-to-interference ratio (SIR) enhancement. Consequently,
20   CDMA network capacities could be significantly increased. In most environments, such as rural, suburban, and urban, transmission sources that are close to a base station create rich, wide angular spread multipath (with continuous distribution in time
25   and space). This will "push" the array response vector away from the array manifold, i.e., increase the Euclidean distance between the array response vector and the array manifold. Distant mobile units, however, provide for a more discrete distribution in time and
30   space, i.e., time distinguishable signal paths have smaller angular spreads. Since most subscribers are on the cell edges (which are the most problematic for capacity), array manifold assisted estimation (MAE), which estimates the array weight coefficients vector by
35   looking for the array manifold closest to the measured array response vector, becomes very practical.

Although the usefulness of array manifold knowledge
degrades as the mobile unit approaches the base
station, the SIR significantly improves for the entire
cell.

5   CDMA demodulators include a Time of Arrival (TOA)
searching mechanism and a plurality of demodulating
channels based on fast Hadamard transformers (FHT) for
IS-95 based systems (such as, e.g., discussed in Andrew
J. Viterbi, *CDMA Principles of Spread Spectrum*
10   *Communications*) or PSK demodulating channels for other
systems (e.g., cdma2000, W-CDMA, and UTRAN). Each
demodulating channel is generally connected to a
selected antenna and tuned to a TOA determined by the
searching mechanism. The products of all demodulating
15   channels are added together (coherent or non-coherent
combining), in compromise between performance and
complexity. Coherent combining requires determination
of relationships among all elements to be combined to
ensure maximum constructive combining (weight vector).
20   Non-coherent combining is done by squaring all combined
elements, thereby eliminating potential destructive
combining through elimination of phase between the
combined elements. Non-coherent combining is simpler
and easier to implement, but less efficient, while
25   coherent combining is potentially more efficient but
requires a complex search. When a sufficient (i.e.,
discriminable) TOA spread exists (for CDMA, it is the
inverse of the chip rate, and for IS-95, it is chip
duration, i.e., 800 msec), the plurality of
30   demodulating channels provides for time diversity that
can enhance the standard spatial diversity employed by
most cellular base stations.

As mentioned above, products from a plurality of
demodulating channels can be linearly combined by
35   calculating the weight vector as part of spatial
processing to enhance system performance. The signal-

to-noise improvement can reach 10*logM, where M is the
number of antenna elements.

5   Efficient signal combining in the antenna array
depends, in part, on the ability to estimate the weight
vector (coefficients) used for combining:

$$P = W^T \cdot V$$

where P is a scalar value describing the result of the
10   combining process, W is the weight vector, and V is the
array response vector.

Fast estimation of the combining coefficients is
essential to get close to this goal of estimating the
weight vector during the channel coherency period,
15   which is less than 1/4 of the inverse of the Doppler
rate. A simple approach is possible to enhance a CDMA
demodulation process in fast fading conditions using an
antenna array to estimate a multipath profile in fading
channels and processing the data for beam formation.
20   By exploiting the array manifold in the estimation of
array coefficients in both reverse and forward links,
the CDMA demodulation process can be enhanced in fast
fading conditions using an antenna array to estimate a
multipath profile in fading channels and processing the
25   data for beam formation.

The multipath profile can be defined as a two-
dimensional distribution function of Multipath Power
vs. AOA and TOA. When a signal arrives at the antenna
array, the antenna outputs are collected into a single
30   vector called the array response vector. A collection
of array response vectors created by stepping the
arrival angle of the signal (in two or three-
dimensional space) produces the array manifold. Every
antenna array can be characterized by an array
35   manifold. The array manifold is a trace within an M

dimensional vector space, where M is the number of
antenna elements, as is discussed in above.

In a non-multipath situation (i.e., an ideal wave
front), the array response vector is "touching" a point
5   on array manifold, i.e., the Euclidean distance is
null. When multipath is present, the array response
vector is a linear combination of all arriving
multipath wave fronts. In this case, the array
response vector is "getting away" from the array
10   manifold, i.e., the Euclidean distance is increasing.
The distance between the array manifold and the array
response vector statistically increases as a function
of multipath level, multipath angle spread, and
interference power. Interference includes the combined
15   sum of thermal noise and other incoming transmissions.
As for any case with a considerable number of
contributing random elements, the distance from the
array manifold to the array response vector is assumed
to have a Gaussian distribution, where the mean lies on
20   the array manifold itself and its variance is related
to the above mentioned elements. As the Euclidean
distance increases, the angular spread increases.
Assuming thermal noise and other transmission
interference can be substantially reduced by
25   integration and de-spreading (CDMA), the major factors
causing large separations between the array response
vector and array manifold are multipath level and
angular spread.

The spatial correlator operation can be described
30   by the following operation:

$$\Omega = V^H \cdot A$$

where V is the array response vector (H denotes
35   Hermitian), and A is the array manifold matrix (with
columns indexed by θ). Each row in A represents one

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element of the array manifold, and each column in A
represents one angle in the array manifold. The result
of the spatial correlator operation is a vector Ω of
values with magnitudes corresponding to the level of
5   correlation between the array response vector and the
array manifold for all given possible angles (array
manifold index, θ). Sorting the magnitude of the
largest Ω element means selecting the best fit
achievable (the theoretical maximum happens when the
10   weight vector W lies on the array manifold), i.e.,
having a point touching on the array manifold. In the
case of angular spread that exceeds the array beam
width (array beam width is a subject well known in the
art), the above process can be described as a moving
15   beam within an angular sector containing Rayleigh faded
signal sources (their combined power is constant),
searching for the maximum value at a given instance.
The larger the sector, the larger the sampling
population, leading to increasing the probability of
20   finding a large power value.

      Since all the above operations are linear, both
the relative amplitude and phase of the incoming signal
are preserved (Ω at the selected θ). Hence, this
process can be utilized for both non-coherent
25   demodulation (e.g., M-ary) and coherent phase
demodulation (i.e., PSK) schemes. In pilot-assisted or
coherent demodulation, the estimation of the relative
phase of each signal path (RAKE finger) can be quicker
and hence more accurate in fast fading environments,
30   resulting in better demodulation efficiency and more
accurate coherent fingers combining at the RAKE
receiver.

      In order to construct the multipath profile,
spatial processing is used to estimate the AOA values.
35   Fig. 14 shows a possible embodiment of a 2D CDMA
demodulator for non-coherent modulation transmission

(IS-95 reverse link) to estimate signal AOA values.   In
Fig. 14, a single "finger" (demodulation channel) of a
Manifold Assisted Spatial Demodulator for an IS-95
based system is shown.   This type of demodulator

5   utilizes the knowledge of the array manifold (as
created in clean environment, i.e., no scattering
sources) to enhance the demodulation process.   In our
case, the incoming array response vector is cross-
correlated against the array manifold matrix to provide

10   for a "magnifying glass" effect.   The system "looks" at
the signal only after the spatial correlation takes
place (magnifying glass) since only then may the signal
to noise ratio be sufficient to make any decision.   The
full MAD implementation includes a plurality of MAD

15   "fingers" (at least two, to allow for minimum time
diversity).   The MAD finger described can perform both
time search and demodulation.

        I and Q components of the signal are fed from an
antenna array with M elements.   The M antenna element

20   outputs are down-converted to a base band frequency and
digitized.   The M signals are then de-spread (as
described above or in Andrew J. Viterbi, *CDMA
Principles Of Spread Spectrum Communications*, which is
incorporated by reference in its entirety) along M

25   parallel correlation channels as each signal is
multiplied by the appropriate long and short codes from
a code generator 1405.   After de-spreading, the signals
are input into a bank of fast Hadamard transformers
(FHTs) 1410. The complex-valued outputs of the M FHTs

30   are then fed to M multiplexers 1415 to multiplex the
outputs into 64 (for IS-95) possible array response
vectors (per possible symbol) and fed one by one into a
spatial correlator 1420.   The spatial correlator
executes the spatial correlator operation described

35   above for each of the 64 candidate array response
vectors.   Each array response vector is cross-

correlated in the spatial correlator 1420 according to
the spatial operation described above with the 256
vectors in the array manifold, although other numbers
of vectors are also possible. The number of potential
5   complex multiply-and-accumulate (MAC) operations
required per symbol (assuming 256 possible angles and
M=16 antenna elements) is

NN = M*64*ANGLE_RANGE = 16*64*256 = 262,100.
10

This corresponds to $262,100*5000 = 1.311*10^{10}$ MAC
operations per second, where the rate is 5000 Hz since
duration of an IS-95 symbol is 200 msec. The estimated
AOAs output from the spatial correlator can then be
15   used for further processing, as described below.

        As disclosed earlier, if a sufficient number of
antenna elements is used (i.e., 6 or more) the array
manifold can be represented with very low resolution or
a small number of bits without losing too much in the
20   magnitude of Ω. Reducing the number of bits allows for
simple ASIC implementation and less required processing
speed since no real multipliers are required and memory
size requirements are small. Hence, the above process
becomes realizable within medium-sized ASICs.

25   In another embodiment, shown in Fig. 14A, a "Max
Absolute Value Sorter" 1425 selects the maximum value
in the matrix Ω (AOA and Walsh Symbol index) resulting
from the spatial correlation operation; which is a 64 x
256 size matrix (M is 64, and number of angle steps for
30   the manifold calibration table is 256). The spatial
correlation operation is repeated several times (the
number of times depends on the coherency time
available, i.e., Doppler period divided by a number
ranging from 5 to 10.) The resultant population of AOA
35   values is averaged to determine the column in the
manifold matrix, which represents a vector in the array

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manifold. This column is used as the weight vector for
the next incoming Walsh symbol's weight vector. This
"next symbol" produces a matrix containing 64 possible
array response vectors, which are each multiplied by
5   the above selected weight vector to produce again M-ary
values. The rest of the process is well known, such as
described in Andrew J. Viterbi, *CDMA Principles of
Spread Spectrum Communications*, at page 100, Figure
4.7.

10   The FHTs of Figs. 14 and 14A can be replaced with
standard complex accumulators 1505 (as shown in Fig. 15
for the demodulator of Fig. 14), such as those
discussed in above-referenced U.S. Pat. Provisional
App. Serial No. 60/077,979, entitled "Capacity
15   Enhancement for W-CDMA Systems", if a pilot signal (or
known continuous training sequence) is embedded in the
transmitted signal for coherent demodulation. The
coherent demodulator or AOA estimator could be
implemented in a cdma2000 type system as described in
20   "The cdma2000 ITU-R RTT Candidate Submission produced
by TR45.5 (TIA)" or in any other embedded continuous
pilot or training sequence assisted demodulation
scheme.

           The results of the de-spreading channels are
25   grouped together to form an M-valued array response
vector, and a search is performed only for a single
possible symbol, instead of 64 possible symbols for
non-coherent estimation, in the embedded pilot signal
or training sequence. In this case the computation
30   load is significantly lighter since the data symbol is
known. Thus, the number of potential MAC operations
is, for 16 antenna elements or a 16-valued array
response vector,

35     $NN = M*ANGLE\_RANGE = 16*256 = 4096$

This corresponds to 4096*Computation Rate = 4096*10000 = $4.096*10^7$ MAC operations per second.

Fig. 16 shows an embodiment of a generalized CDMA AOA/MAG (magnitude of Ω) estimator. In Fig. 16, a

5 single "finger" (demodulation channel) of a Manifold Assisted Spatial Demodulator suitable for IS-95 (A,B, or C), cdma2000, and W-CDMA/UTRAN proposals is shown. In this embodiment, the de-spreading mechanism can follow W-CDMA (NTT/DOCOMO) and UTRAN (ETSI/SMG)

10 proposals to ITU, which are proposals in response to the ITU 3rd Generation cellular IMT-2000 initiative. The main difference relative to the current IS-95 (A and B) standard is the existence of a pilot signal in the reverse link. The IS-95C and cdma2000 proposal

15 employ a continuous pilot signal, while W-CDMA employs evenly spaced short bursts of a pilot signal. Details of the reverse link structure are given in the CDG cdma2000 and ETSI/SMG & NTT DOCOMO W-CDMA UTRAN/ARIB proposal submitted to the ITU on June 1998, which is

20 incorporated by reference in its entirety.

I and Q components of a received signal are fed into an antenna array with M elements. The M antenna element outputs are down-converted to a base band frequency and digitized. The M signals are then de-

25 spread along M parallel channels, such as above. For W-CDMA, the de-spreading blocks 1605 can be bypassed for the initial mobile station access phase. When W-CDMA's mobile station's timing has been established, the de-spreading blocks can be used for de-scrambling.

30 For cdma2000, the de-spreaders 1605 are used as suggested in the CDG proposal (accommodating long and short codes). The proposal to the ITU, incorporated above, provides additional details about the de-spreading.

35 The next stage, at the output of the de-spreader blocks 1605, comprises a bank of M matched filters 1610

for W-CDMA or IS-95 or a bank of M accumulators 1610
for cdma2000. The matched filters are gated to allow
for a non-continuous pilot signal as proposed by W-CDMA
and UTRAN, i.e., utilizing matched filters for the 256
5   bit sequence and scrambling code combined, as suggested
by UTRAN and W-CDMA. The matched filters correlate the
incoming signal sequence with a pre-stored sequence.
The outputs of the matched filters are fed into a
spatial correlator 1615 for determining the best fit in
10   the array manifold matrix using the operation

$$\Omega = V^H \cdot A$$

as explained above.
15   It should be noted that the time of signal arrival can
vary and needs to be tracked. The ability to determine
when the matched filter produces a response to the
training sequence may be limited due to low signal to
noise ratio conditions, requiring repetitive hypothesis
20   (i.e., changing the sampling time). This can be done
only after the spatial correlation, which requires the
spatial correlation operation to be very fast.
       For W-CDMA, a new array response vector group is
generated every 0.625 msec. In the case of time
25   distinguishable multipath, several array response
vectors may be generated sequentially in this time
frame. The time separation depends on the multipath
TOA spread. In rich multipath environments, there may
be up to L distinguishable multipath elements (e.g.,
30   L=3). Since the exact timing and phase of the training
sequence cannot be determined before the spatial
correlation block ("magnifying glass"), time varying
sampling of the matched filter outputs is required
(time search) allowing incremental time hypotheses.
35   Each hypothesis requires a spatial correlation process,
hence, the spatial correlation process determines the

search time to acquire a mobile station. The current
spatial correlator design allows up to 200,000 spatial
correlation operations per second. For fast fading
conditions, the rate of estimation updates for time

5   hypotheses may reach 500,000 times per second (1000
times faster than the maximum Doppler rate). In this
case, the number of MAC operations per second is
$4096*500000*L = 20.48*10^8*L$ for a 16-element antenna or
a 16-valued array response vector and 256 possible

10 angles. If L=3, the number of MAC operations per
second is $6.144*10^9$. Utilizing the low bit count
algorithm described above, this rate is very feasible
using current ASIC technology.

For cdma2000, the outputs of the de-spreaders are

15 grouped together in the accumulators to form an M-
valued array response vector, and a search is performed
for a single possible symbol in the embedded continuous
pilot signal. The time search is similar to searches
for IS-95 based systems. Supporting different TOA

20 multipaths can be performed by either utilizing a
single searcher (using the same spatial correlator or
duplicating the "finger" described in Fig. 15). Figs.
14, 14A, 15, and 16 describe a mechanism to deal with a
single time of arrival path. In the case of multiple

25 time of arrival paths, multiple modules are suggested.
Rather than just adding additional modules, different
embodiments can share some common circuitry to reduce
the overall circuit size and cost.

Once estimated AOA data is available, such as from

30 a spatial correlator in Figs. 14, 14A, 15 or 16, the
data is processed to enhance receiver performance,
which includes three components: 1)beams with
sufficient gain must be formed toward the incoming
signal, 2) spatial diversity must be provided, and 3)

35 the downlink beam must be constructed.

The data from the demodulator (from Figs. 14, 14A,
15, or 16 above) is collected to form AOA histograms.
Since the mobile unit provides a changing wave front
(wave front is a linear combination of many incoming
5   wave fronts from many scatterers), a continuous
accumulation of AOA samples allows an AOA histogram to
build up. This histogram will have "peaks" in the
direction of the main scatterers and a distribution
that follows the angular spread of the transmission
10   source. A significant advantage of the AOA histogram
is the ability to distinguish the peaks even if the
transmission is non-continuous (as for IS-95 based CDMA
systems). After determining the AOA histogram peaks
and variances, beams can be formed in the directions
15   associated with the peaks and with widths that follow
the histogram variances. In the case of a single AOA
peak, the system can form multiple beams offset in a
direction toward the main direction. If the array is
large enough, it can be shown that the power from the
20   signals derived from the various beams have a low
correlation. This correlation is derived from the
inner product of different columns in the array
manifold matrix. Exploiting the fact that most CDMA
systems use some implementation of RAKE combining, each
25   RAKE channel can be connected to a different beam.
This arrangement achieves the first two components
mentioned above: gain and diversity.
      Another feature of the AOA histogram processing is
the ability to estimate the downlink beam, which is the
30   third component mentioned above. Although in FDD
systems there is a difference between reverse and
forward link frequencies, there is also a good
statistical relation. Hence, the forward link beam is
formed with a direction and width following the
35   histogram distribution. In the new generation systems,
pilot signals can be made available on the forward

link, and hence, no specific effort is needed to
accommodate phase coherency between the system's main
pilot and the forward traffic channel. For IS-95 based
systems, such as described above, matched phase beam
5   synthesis is utilized on the forward link.

        In another embodiment, shown in Fig. 16A, a phase
rotator 1620 and an inner product multiplier 1625,
which can be integrated with the demodulator of Fig.
16, further process the result from the spatial
10 correlator 1615. In both IS-95C/cdma2000 and W-CDMA
cases, the array response vector (or array response
vector group) can be produced by integrating data over
time that is limited by the Doppler rate (or a small
fraction thereof) to minimize lagging errors. The
15 weight vector and carrier phase (PSK) need to be
estimated for demodulation and beam forming. The time
to do so is limited by the coherency period which is
equal to a small fraction of the Doppler period. The
spatial correlator, which enhances the signal to noise
20 ratio, allows a faster determination of these values.
The result of the spatial correlation is a pointer to
the best-fit column in the array manifold calibration
table. The index of the maximum of the resulting
correlation matrix Ω is the pointer. The phase of the
25 maximum value (selected as part of the spatial
correlator process) is the carrier rotation phase. The
selected column (W) in the array manifold matrix, which
contains the maximum valued element of the matrix, from
the spatial correlator 1615 is fed into phase rotator
30 1620 to shift column W. The shifting is done by
multiplying W by $e^{j\emptyset}$, where the phase $\emptyset$ is the argument
of the maximum value selected from the spatial
correlator resultant vector Ω:

35                          $$\Omega = V^H \cdot A$$

50

The shifted column, W', is then fed into a
multipliers bank module 1630 of inner product
multiplier 1625. The inner product multiplier, which
also includes an adder circuit 1635, performs a typical
5   beam forming operation with a difference being that the
weight vector W' has been phase adjusted to the
incoming signal array response vector to maximize PSK
(phase shift keying) demodulation results. The
multipliers bank module performs the following
10   operation for beam forming:

Conjugate(Selected column from A shifted by Ø
$(W*e^{j\varnothing}))^T$ * array response vector.

15   The efficiency of this demodulation process depends on
various factors, such as the accuracy of the array
manifold calibration matrix selection (i.e., AOA
estimation), the accuracy of the rotation phase
estimation, the amount of angular spread, and the SIR.
20   Figs. 17 and 18 show a performance comparison between a
standard two element diversity array and the MAD system
described above for same signal fading conditions.
Fig. 17 shows results using a QPSK (Quadrature Phase
Shift Keying) MAD in random fading. Fig. 18 shows the
25   same results of the same fading conditions as in Fig.
17, but with a standard QPSK demodulator. The
simulation results show an average improvement of
approximately 6 to 8 dB for the MAD based system.
For a typical base station, the number of
30   simultaneous mobile station sessions may reach 100 or
more, which would require multiplying the number of MAC
operations per second mentioned above by 100 and more.
The ability to reduce the number of bits in the process
allows for practical ASIC implementation. Each voice
35   or data channel is equipped with a spatial correlator
that executes the above-described operation. The

result is a spatially enhanced demodulator, i.e., for
each received symbol, the system searches for the best
way to coherently combine the output of all the antenna
ports.

5   The effectiveness of this spatially enhanced
demodulation grows as the mobile unit gets farther away
from the base station because the greater the distance,
the smaller the multipath angular spread, and hence,
the array response vector lies closer to the array
10   manifold. As the array response vector gets nearer to
the array manifold, the accuracy of the signal AOA and
magnitude estimation increases due to the smaller
multipath angular spread. Assuming a uniform
distribution of mobile units across the network, most
15   mobile units are at the outer regions of the cellular
cell. In addition, the farther the mobile unit is from
the base station, the harder the communication is to
maintain. Since distant subscribers are harder to
maintain communication with, solutions directed to
20   distant subscribers are a higher priority. Thus,
accuracy degradation and reduction of demodulator
efficiency at close proximity to base station is
tolerable.

          Fig. 19 presents an embodiment illustrating a
25   training sequence convolver, which may be used instead
of the de-spreader 102 and FHT 103 in some wireless
standards. A data register 1902 is a first-in-first-
out (FIFO) unit with a word bandwidth that is matched
to the receiver I and Q output width. The I and Q
30   samples are shifted through the data register 1902 in
two's complement format. XOR gates are used to compare
the most significant bit of I and the most significant
bit of Q with bits of a training sequence stored in a
training sequence register 1903. The resultant XOR
35   output are fed to an adder 1901 and used to determine
whether to add or subtract each I and Q sample in the

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data register. The output of the adder is updated for
every sample cycle and compared against a threshold in
a magnitude threshold detector 1904. When the
threshold is exceeded, the I and Q values are
5   registered as a component of the signal response vector
that is then sent to the spatial correlator explained
above.
       Fig. 20 presents an embodiment of the invention
that includes both searching and tracking functions (in
10   angle and time). The addition of angular tracking
increases the ability of the system to efficiently
direct the receiving beams at all times. A searcher
2000 acquires new multipath parts as before while a
tracker 2003 tracks them. The principle of operation
15   of this embodiment is very similar to the embodiment
described in Fig. 1. The main difference relative to
Fig. 1 is the addition of the tracker 2003. The N
receiver outputs are fed in parallel to beamformer
2012. The controller 2001 downloads to the beamformer
20   2012 not just one, but two beam-forming information
sets for each signal part to be tracked. The two sets
correspond to two adjacent columns in the calibration
matrix. This allows the beamformer to continuously
"toggle" between two angularly adjacent beams.
25   The beamformer output is fed into an "Early/late
gate" module 2013 known in the art. The result of the
combined "toggling" beamformer and the "Early/late
gate" is in the form of four level values corresponding
to: left beam/early time, right beam/early time, left
30   beam/late time and left beam/late time. Since the
tracker is designed to track four multipath parts
simultaneously, the results are reported to the
controller through a multiplexer 2015. The controller
2001 directs the beamformer and the "Early/late gate"
35   to balance all the four values above the same level by
exchanging the beamformer coefficients and

advancing/delaying the gate's clock. Angular tracking
is achieved by equalizing the right and left associated
results while the time tracking is achieved by
equalizing the early late associated values. This
5   embodiment assures sufficient integration for reliable
tracking. The sets of coefficients are entirely
replaced when the searcher finds a multipath part that
generates significant higher level outputs than the
ones tracked. In this embodiment, each channel is
10 assigned its own downlink beamformer 2030. Note also
that this embodiment supports an individual beam for
each active channel.
    Fig. 21 presents an overview of a base station
that employs channel estimators/trackers/beamformers
15 described in Fig. 20. The antenna array 2100 is
coupled to a set of receivers 2101 which are all driven
by common local oscillator 2104, as in Fig. 1. The
receiver outputs are placed on a data bus 2110 to feed
a plurality of channel estimators/trackers/beamformers
20 2105, each of which provides a BTS channel element 2106
with a plurality of signal parts. Element 2106 can be
a RAKE receiver/data transmitter of IS-95. The channel
elements are feeding downlink data to the channel
estimator/tracker/beamformers, which feed beam-formed
25 data to summation unit 2107. The summation unit
outputs summed beam-formed data to the BTS transmitters
2109 that are driven by common local oscillator 2108.
The transmitter outputs are radiated through
transmitting antenna array 2111.
30   The above embodiment of the downlink requires
additional "pilots" when applied to a CDMA IS-95 base
station. This may require some changes in the network
control and network pilots allocation design. The
following embodiment alleviates this requirement by
35 distributing the overhead channels (pilot, paging and
synchronization) through a wide beam while the traffic

channels are individually transmitted through narrow
beams directed at the associated mobile units. This
approach does not change the conventional BTS softer
hand-off profile, hence it does not require any changes
5   in the network architecture.
     This proposed arrangement is facilitated by
careful array beam synthesis techniques that are well
known in the art. In particular, the beams are
constructed to be phase matched in the mobile unit's
10   scattering region. The beams' coefficients are
calculated to achieve identical wave-fronts between the
pilot and the traffic signals, hence, allowing the
current IS-95 coherent demodulation at the mobile unit.
This "beam matching" is facilitated using beam
15   synthesis based on a minimum root mean square approach.
This approach allows for +/- 10 degrees phase matching
down to -10 dB points, which is sufficient not to
degrade the performance of the coherent demodulator at
the mobile unit.
20   The coefficients of the individual downlink beams
are set as follows: the overhead data (pilot, synch
and paging) are transmitted through fixed, relatively
wide beams. The downlink traffic data beams are set to
match the line of bearing as measured by the uplink
25   channel estimator with sufficient width margin to
compensate for bearing error (due to lack of
correlation between up and down links). It should be
noted that even relatively wide downlink traffic beams
will provide for significant capacity increase.
30   Since the angular spread is getting larger as the
distance to the base station decreases, the narrow beam
width is estimated based on the estimated distance from
the BTS. This distance is derived from the time delay
as measured by a beam director.
35   Since the above approach is based on the
statistical profile of the scattering region (various

55

scattering models are considered), the system must
provide for exceptions: at first, the allocated
traffic narrow beam is made wider than needed, and as
done with the forward power control, it is gradually

5   narrowed based on the frame erasure rate (analogous to
bit error rate) that is reported on the uplink. In
case the frame erasure rate increases, the traffic beam
is widened accordingly. This mechanism will also
compensate for situations where the uplink angle of

10 arrival (AOA) is very different from the downlink AOA.
        While the above embodiments describe utilizing the
present invention for current CDMA communication
systems, the concepts of the present invention can also
be used in wide band CDMA (W-CDMA) communication

15 systems to increase system capacity. More
specifically, a W-CDMA system utilizes a plurality of
antennas arranged in wide aperture array and digital
signal processing to estimate multipath angle of
arrival (AOA) and time of arrival (TOA), thereby

20 allowing the assignment of multi-antenna beams towards
the incoming signal parts and the assignment of
adjustable downlink beams to increase system capacity.
Even though the specifications for W-CDMA are not yet
clearly defined, there are already specific principles,

25 such as the existence of a pilot signal in the uplink,
that are accepted (e.g., part of IS-665 and J-STD-015)
with regard to W-CDMA that can be exploited to provide
for efficient adaptive antenna array technology
implementation.

30   To provide for capacity enhancement of W-CDMA
communication systems, the following features are
proposed: uplink channel estimation, beam-forming for
uplink to provide for enhanced antenna array gain,
spatial directivity and fading mitigation (through

35 diversity), and beam-forming for downlink for enhanced

array gain and spatial directivity, which will be
discussed in order below.

    As previously discussed, signal de-spreading and
fast Hadamard transformers (FHT) can be used to
5   estimate the array response vector (electrical
amplitude and phase for all array elements) of an IS-95
based CDMA signal to provide for multipath AOA values
determination through spatial correlation. However,
the existence of a pilot in the uplink for a W-CDMA
10 system, which is a basic difference between W-CDMA and
IS-95 based CDMA systems, can be utilized to determine
the array response vector and channel impulse response
(CIR) for determining AOA and TOA values for uplink
channel estimation. Fig. 22 shows a possible
15 implementation of W-CDMA uplink traffic channel, which
is defined and described in the CDG cdma2000 proposal
submitted to the ITU on June 1998, incorporated above.

    The presence of pilot data in the uplink allows
the estimation of the array response vector as for
20 training sequences. A regular W-CDMA receiver
synchronizes its demodulator to the incoming W-CDMA
signal by hypothesizing on the periods of the pilot
data streams. Each hypothesis consists of accumulating
k number of incoming signal samples and multiplying the
25 samples by k number of internally generated replicas of
pilot samples (i.e., the inner product between the
incoming signal and the generated replica signal). The
pilot's replica sequence is delayed for each subsequent
hypothesis and the correlation process repeats. When
30 the replica of the pilot is synchronized with the
incoming signal, the resulting I and Q magnitudes are
maximized to indicate "lock" conditions. Continuing
the above accumulation process, accurate pilot, and
hence, carrier phase can be determined.

35   Since the pilot part to the incoming signal exists
all the time, the integration period is limited only by

<center>57</center>

the movement of the mobile unit (Doppler shifts) and
the inaccuracy of the carrier frequency used at the
receiver demodulator. Because for a typical mobile
unit speed, the doppler shift is less than 100 Hz, and
5   the frequency error is usually in the order of hundred
hertz, the integration period could span over several
milliseconds, which is typically far longer than a
symbol duration. This mechanism is similar to
demodulation in the IS-95 downlink, described above.
10   Using the pilot correlation process described
above, the carrier relative electrical phase can be
estimated using a phase estimator 2300 as shown in Fig.
23. The incoming signal from the mobile unit is
divided into two branches by a power divider 2301 and
15   multiplied by a signal generated from a quadrature RF
signal generator 2302 to produce I and Q signals.
After each signal is passed through a baseband filter
2303 and analog-to-digital (A/D) converter 2304 for
baseband filtering and digitizing, respectively, the I
20   and Q sample streams are fed into multiply-and-
accumulate (MAC) and scale circuitry 2305. A delayed
pilot code sequence, as described above, from a pilot
sequence generator with variable delay circuit 2306 is
multiplied with the I and Q sample streams from A/D
25   converter 2304, summed, and scaled to produce SUM(I)
and SUM(Q) quantities, which represent a single element
within the array response vector. If the delayed pilot
code sequence differs by more than one chip duration
from the incoming signal sequence, the SUM(I) and
30   SUM(Q) values are small (following the auto-correlation
function of the pilot sequence). Hence, the phase
estimator can be used as a Channel Impulse Response
(CIR) estimator by varying the delay values from the
pilot sequence generator 2306 over the expected range
35   of incoming signal TOAs.

The CIR is essential to accurately determine the
signal multipath AOA and TOA values. The existence of
pilot signal in the uplink can be used to determine the
CIR. As discussed above, SUM(I) and SUM(Q) output
5   magnitudes depend on the time difference between the
incoming signal and the internally generated pilot
replica sequence. A conventional searcher (in a RAKE
receiver) varies the delay of the internally generated
pilot sequence while evaluating the square of the sum
10   of SUM(I) and SUM(Q) (i.e., $[SUM(I)+SUM(Q)]^2$] to measure
the CIR.

Fig. 24 shows a system (W-CDMA BeamDirector™) for
enhancing the normal search process utilizing spatial
correlation as described earlier. Signals from antenna
15   elements in the receiving antenna array are processed
in dual banks of N phase estimators, such as phase
estimators 2300 in Fig. 23, to generate SUM(I) and
SUM(Q) signal components. Each set of N SUM(I) and
SUM(Q) components is correlated with an antenna array
20   calibration matrix by a spatial correlator 2400 to
produce a correlation matrix that represents the
correlations of the signal received at the antenna
array with both a set of predetermined directions and a
set of predetermined symbols, such as discussed above
25   with Fig. 4. The results of the spatial correlator are
read by a controller 2401 to produce the CIR data (both
magnitude and AOA data). Fig. 25 shows an example of
the CIR data as functions of time of arrival.

The controller 2401 analyzes the CIR data to
30   determine which TOA values are to be used by a "House
Call" section 2402 containing the first bank of phase
estimators and a spatial correlator. The "House Call"
section 2402 is very similar to a searching section
2403, which contains the other bank of phase estimators
35   and a spatial corrrelator. However, the "House Call"
section dwells on the TOA values that were determined

from the CIR data as multipath TOA values. This
mechanism allows for a high success-to-attempt ratio in
measuring AOA data for the incoming multipath parts.

The angle/time of arrival estimation described
5   above allows for both single scattering zone and multi-
scattering zones handling. Angular spread can be
determined in real time by way of histogram processing
of angle of arrival samples. When fading is produced
by a large scattering zone, the angle of arrival
10   results (AOA samples) are distributed with large
variation (can be estimated by the variance of AOA
results). The main AOA, however, can be estimated by
the histogram center of gravity. The histogram center
of gravity is determined by "smoothing" the histogram
15   through a low pass filter (e.g., Hamming, Raised
Cosine, etc.) and finding the maximum point of the
"smoothed" histogram. The multipath scattering area
size can then be estimated by comparing the "smoothed"
histogram peak value to the histogram data
20   distribution. When more than one scattering zone
exists, thereby causing multiple "peaks" in the CIR
data, a separate histogram process is performed for
each significant "peak" associated with TOA value in
the CIR.
25   The estimated AOA values along with scattering
zone sizes (sectorial angle) are then used to determine
the coefficients of the uplink beamformers bank 2404
that feeds the uplink RAKE receiver. Since the number
of RAKE receiver "fingers" is limited, the assignment
30   of uplink beams is optimized to maximize the RAKE
combining efficiency. For example, if only a single
scattering zone is identified, all beams are arranged
to evenly cover the identified scattering zone. If
multiple scattering zones are identified, the beams are
35   allocated to assure first that all distinct scattering
zones are covered, and then the remaining available

beams are added to provide diversity within the more
dominant scattering zones. Coefficients for the
downlink beamformer 2405 that go to the transmit
antenna array can also be determined using CIR data
5   from the controller following the same downlink
principles discussed above. The beam width is
determined from the uplink multipath distribution, and
the beam coefficients are set to assure illumination of
the scattering zone as determined by the multipath
10   distribution.

As is evident from the various embodiments
illustrated above, the present invention encompasses
within its scope many variations. Those skilled in the
art will appreciate that additional modifications may
15   also be made to the above embodiments without departing
from the scope of the invention. Accordingly, the true
scope of the present invention should not be construed
as limited by the details provided above for the
purposes of illustration, but should be determined from
20 the following claims and their legal equivalents.

Claims:

1.   A method for wireless communication
comprising:

a)   transmitting from a mobile unit a code
modulated signal obtained by modulating original
symbols by a predetermined pseudo-noise sequence,
wherein the original symbols represent an original
information signal;

b)   receiving at a base station antenna array
N complex valued signal sequences received in
parallel from N corresponding antenna elements;

c)   correlating in parallel each of the N
signal sequences with the pseudo-noise sequence to
select N received signals comprising N received
symbols corresponding to a common one of the
original symbols;

d)   transforming in parallel the N received
symbols to obtain N complex-valued transformer
outputs;

e)   correlating collectively the N
transformer outputs with a set of complex array
calibration vectors to obtain spatial information,
wherein each array calibration vector represents a
response of the antenna array to a calibration
signal originating in a predetermined direction
relative to the base station;

f)   repeating steps (b),(c),(d),(e) to obtain
spatial information about multiple signal
components;

g)   spatially filtering a subsequent set of N
complex valued signal sequences in accordance with
the spatial information about multiple signal
components; and

h)   demodulating the spatially filtered
subsequent set to obtain a symbol from the
original information signal.

2.   The method of claim 1 further comprising
tracking time and angle information of the multiple
signal components.

5

3.   The method of claim 1 wherein the original
symbols are selected from a symbol alphabet comprising
not more than 64 symbols.

10   4.   The method of claim 1 wherein each of the N
transformer outputs comprises a vector having M complex
valued components representing correlations between a
received symbol and M symbols of a symbol alphabet.

15   5.   The method of claim 1 wherein the calibration
vectors comprise complex-valued components having 1-
bit-plus-sign real part and 1-bit-plus-sign imaginary
part, and wherein the correlating comprises computing
via addition only a vector dot product between the
20   calibration vectors and the N transformer outputs.

6.   The method of claim 1 wherein the correlating
yields spatial information about multiple signal
components having a time spread less than one chip.

25

7.   The method of claim 1 further comprising
spatially filtering a downlink information signal in
accordance with the spatial information about the
multiple signal components, and transmitting the
30   spatially filtered downlink information signal from the
antenna array to the mobile unit.

8.   The method of claim 7 wherein the spatially
filtering comprises assigning the mobile unit to a
35   calculated beam and generating the beam.

9.   In a wireless communication system comprising a mobile unit and a base station having an N-element antenna array, a method for efficiently determining at the base station a spatial channel of the mobile unit,

5   the method comprising:

a)   calculating a transform of a symbol as received from a first antenna of the antenna array, wherein the calculation produces a first M-dimensional vector having complex-valued

10   components, where M is a number of predetermined symbols in a symbol alphabet;

b)   performing step (a) simultaneously and in parallel for the symbol as received from N-1 additional antennas in the array, thereby

15   producing a matrix B containing N row vectors of dimension M;

c)   calculating the matrix product $C=A^H B$, where each of L columns of the matrix A is an N-dimensional vector containing a response of the N

20   antenna array in one of L predetermined directions relative to the array; and

d)   determining from the matrix C a spatial direction of a signal part originating from the mobile.

25

10.   The method of claim 9 wherein $M < 65$.

11.   The method of claim 9 wherein the matrix A has complex valued elements having 1-bit-plus-sign real

30   part and 1-bit-plus-sign imaginary part, whereby the matrix product calculation is efficiently performed.

12.   The method of claim 9 further comprising determining from the matrix C an additional spatial

35   direction of an small time separated signal part originating from the mobile.

13. A method for wireless communication comprising:

transmitting an information signal from a mobile unit;

receiving the transmitted signal with an array of N antenna elements to yield a set of N received signals;

spatially correlating the N received signals with the contents of an antenna array calibration table to obtain directional information about the mobile unit, wherein the stored calibration table comprises complex-valued elements having 1-bit-plus-sign real part and 1-bit-plus-sign imaginary part, whereby spatial correlation is facilitated; and

spatially filtering subsequent received signals from the mobile unit in accordance with the directional information to obtain corresponding transmitted information signals.

14. The method of claim 13 wherein the receiving comprises digitizing, de-spreading and Hadamard transforming, separately and in parallel, N air signals coupled to the N antennas.

15. The method of claim 13 wherein the correlating comprises calculating vector dot products between the N received signals and columns of the array calibration table having complex-valued elements in the form of a bit-plus-sign real part and a bit-plus-sign imaginary part.

16. The method of claim 13 further comprising assigning the mobile to a calculated downlink beam based on the directional information.

17. The method of claim 16 wherein the calculated beam is selected from among a dynamically adaptive set of overlapping downlink beams of differing angular
5   extent.

18. The method of claim 16 wherein the assigning is further based upon distance information such that close mobiles are assigned to broad beams and distant
10 mobiles are assigned to narrow beams.

19. A method for wireless communication comprising:
        transmitting uplink information signals from
15   a set of mobiles;
        receiving the uplink signals with an array of N antenna elements to yield a set of N received signals;
        processing the N received signals to obtain
20   spatial information about the mobiles;
        calculating downlink beam-forming information based upon the spatial information, wherein the beam-forming information comprises assigning each of the mobiles to one of a set of downlink beams,
25   wherein the set of downlink beams comprises wide beams for nearby mobiles and narrow beams for distant mobiles, and wherein the wide beams overlap the narrow beams; and
        transmitting downlink information signals to
30   the mobiles in accordance with the calculated downlink beam-forming information.

20. The method of claim 19 further comprising tracking the spatial information about the mobiles in
35 angle and time.

21. The method of claim 19 further comprising modifying the properties of the downlink beams based upon the spatial information in order to optimize system performance.

5

22. The method of claim 19 wherein the transmitting is performed in accordance with beam-forming information comprising complex valued elements having 3-bit-plus-sign real part and 3-bit-plus-sign

10 imaginary part.

23. The method of claim 19 wherein the processing comprises Hadamard-transforming the N received signals and correlating the N transformed signals with an array

15 calibration table.

24. The method of claim 23 wherein the array calibration table comprises complex-valued elements having 1-bit-plus-sign real part and 1-bit-plus-sign

20 imaginary part.

25. The method of claim 23 wherein the correlating comprises a matrix multiplication implemented as a complex addition.

25

26. A CDMA base station comprising:
an antenna array comprising N antenna elements;
a set of N receivers coupled to the N antenna
30 elements to produce N incoming signals;
a set of N de-spreaders coupled to the N receivers, wherein the de-spreaders produce from the N incoming signals N de-spread signals corresponding to a single mobile unit;

a set of N symbol transformers coupled to the
N de-spreaders, wherein the transformers produce
complex-valued outputs from the de-spread signals;
a spatial correlator coupled to the N symbol

5 transformers, wherein the correlator correlates
the complex-valued outputs with stored array
calibration data to produce beam-forming
information for multiple signal parts associated
with the mobile unit;

10 a receiving beamformer coupled to the spatial
correlator and to the N receivers, wherein the
receiving beamformer spatially filters the N
incoming signals in accordance with the beam-
forming information; and

15 a RAKE receiver coupled to the receiving
beamformer, wherein the RAKE receiver produces
from the spatially filtered signals an information
signal.

20 27.   The base station of claim 26 further
comprising a transmitting beamformer coupled to the
spatial correlator, wherein the transmitting beamformer
generates spatial beams in accordance with the beam-
forming information.

25

28.   The base station of claim 27 wherein the
spatial beams are selected from a set of calculated
beams comprising narrow beams and overlapping broad
beams, where the narrow beams are phase matched to the

30 overlapping wide beams.

29.   The base station of claim 26 further
comprising a tracker coupled to the spatial correlator
and to the receiving beamformer, wherein the tracker

35 tracks the multiple signal parts and optimizes the
performance of the receiving beamformer.

30.   The base station of claim 26 wherein the
array calibration data comprises complex valued array
response elements represented as bit-plus-sign
5   imaginary parts and bit-plus-sign real parts.

31.   The method of claim 1 wherein the calibration
vectors comprise complex-valued components having 2-
bit-plus-sign real part and 2-bit-plus-sign imaginary
10   part, and wherein the correlating comprises computing
via addition only a vector dot product between the
calibration vectors and the N transformer outputs.

32.   The method of claim 1 further comprising
15   forming multiple narrow beams in a wide aperture
antenna array, wherein said narrow beams cover a
desired scattering zone.

33.   The method of claim 32, wherein the number of
20   the narrow beams is two to four.

34.   The method of claim 32, wherein the width of
the narrow beams ranges from 2 to 3 degrees.

25   35.   The method of claim 1, further comprising
code multiplexing a pilot signal into the code
modulated signal.

36.   The method of claim 35, further comprising
30   corrrelating the pilot signal with delayed pilot
signals generated by the base station.

37.   The method of claim 36, further comprising
forming angle of arrival and time of arrival histograms
35   using the correlation data to form uplink and downlink
beams directed to a desired scattering zone.

38. The method of claim 9, further comprising inserting a sounding signal into transmission and receiving channels.

5

39. The method of claim 38, further comprising multiplying signals from the transmission and receiving channels with the sounding signal to produce a compensation vector.

10

40. The method of claim 39, further comprising adjusting the matrix A using the compensation vector for amplitude and phase compensation.

15

41. The method of claim 1, further comprising calculating the matrix product $\Omega = V^H A$, wherein A is the array manifold matrix and V is the array response vector, and utilizing entries in the matrix Ω to form an angle of arrival histogram.

20

42. The method of claim 41, further comprising using peak and variance information from the histogram to form beams with desired width and direction.

25

43. The method of claim 41, wherein using the matrix Ω at longer communication distances with smaller angular spread.

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FIG. 1

FIG. 2

FIG. 3

FIG. 4

FIG. 5

FIG. 6

FIG. 7

Capacity Increase
Ratio

High
Angular
Spread
Multipath
Ratio/10

Number
of
Narrow
Beams

Ψ

FIG. 8

Capacity Increase
Ratio

Ratio
of High
Angular
Spread
Mobiles/10

Softer
Handoff
Ratio/10

Ψ

FIG. 9

High Angular Multipath Ratio
=0.2 =0.5

Capacity Increase Ratio

$\dfrac{W1}{W2}$ b

Softer
Handoff
Ratio = b*10

b

FIG. 10

Number of Narrow Beams
N=1 N=2 N=3 N=4

Capacity Ratio

Irregularity Variance

FIG. 11

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```
                    ┌─────────┐
                    │  Start  │
                    └─────────┘
```

MS Status Data Base Entries:
- LOB & Qualify Per De-coupled Signal Parts
- TX Power Setting
- MS Multipath Profile
- Approximated MS Distance From BTS
- Pilot Used
— 1200

1205 — Create Two Dimensional Histogram of MS's (Angle & Distance)

1210 — Identify Histogram "Peaks:" (Location & Values)

1215 — Sort MS's By Pilots

1220 — Count The # of MS's Around The M Highest Histogram Peaks (Distance From Peak Determined According To Peak's Value)

1225 — Compare MS Counts (Per Peak) To Closest Beam's Pilot Count

1230 — MS Peak Count Close To Pilot MS Count?

Yes

No

1235 — Shift Closest Beam Towards Selected "Peak"

FIG. 12

1301 — Antenna Array

1302 — TX Antenna

1303 — RF Switch

1304 — Turn Table

1305 — Network Analyzer

1306 — Control & Data Storage Unit

FIG. 13

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FIG. 14

FIG. 14A

FIG. 15

FIG. 16

FIG. 16A

FIG. 17

FIG. 18

FIG. 19

FIG. 20

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FIG. 21

FIG. 22

FIG. 23

FIG. 24

FIG. 25